ISOLATION AND MULTIPLE OUTPUT EXTENSIONS OF A NEW OPTIMUM TOPOLOGY SWITCHING DC-TO-DC CONVERTER

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ABSTRACT

The recently introduced new optimum topology dc-to-dc converter is extended in a simple and elegant manner to provide dc isolation and multiple outputs. In comparison with the single-transistor isolated forward and flyback converters operated under the same conditions, the single-transistor isolated new converter is shown to have equal or lower stress levels on the transistor, diode, and capacitor ripple current, and can utilize an isolation transformer with lower core and copper losses. Measurements of cross- and self-regulation properties of a two-output 45 W test converter are presented.

1. Introduction

A new dc-to-dc converter, introduced at the 1977 PESC [1], was described as having an "optimum topology" configuration because it provides the basic dc-to-dc conversion property with the smallest number of elements that permit both the input and output currents to be nonpulsating. The potential performance, efficiency, and cost benefits to be obtained by use of the new converter were described, and a favorable comparison was made with the conventional buck-boost converter, which has the same dc-to-dc transformation property as does the new converter.

In its original form as described in [1], the new converter is a nonisolated polarity inverting converter. Since many practical applications demand dc isolation, there is strong motivation to extend the new converter configuration to incorporate an isolation transformer. This paper introduces a simple and elegant solution to this problem, in which the original single-transistor converter is augmented merely by a single-ended isolation transformer and an additional capacitance [2].

Similarly simple single-transistor, single-ended transformer-isolated versions of the conventional buck and buck-boost converters are well-known respectively as the "forward" and "flyback" converters. If the new isolated converter is to be viable, its properties must compare favorably with those of comparable forward and flyback converters. In this paper a detailed comparison is made which shows that the new converter has distinct advantages in almost all respects.

This work was supported in part by the Naval Ocean Systems Center, San Diego, California, through MIPR No. N0095377MP09018, and by the International Business Machines Corporation, Kingston, New York.

In particular, the transistor and diode current and voltage stress levels, and capacitor ripple current stress levels are, in most operating conditions, *less* in the new converter than in the forward or flyback converters. If the same isolation transformer core and copper are used in all three, the copper loss also is *less* in the new converter; however, a core of half the area can in fact be used in the new converter, which leads to half the core loss and even lower copper loss than in the forward or the flyback converter.

Once an isolation transformer is introduced, extensions to multiple outputs of various polarities are obvious, and examples are given. Not so obvious is the fact that any or all of the input and output inductors in the multiple-output new converter can be coupled (wound on the same core) with lowered ripple current properties (even zero), as has been described for the original nonisolated converter [3].

Finally, some experimental results on a two-output isolated new converter are presented together with measurements on the cross- and self-regulation properties, which are of concern when the converter is embedded in a feedback loop in which only one output is regulated. This configuration is typical in computer power supplies, among others.

2. The original optimum-topology new converter

The simplest form of the new converter circuit is shown in Fig. 1. Its basic operation and properties have been discussed in [1], and will be only briefly summarized here.

The salient feature of the new converter is that its properties closely approach those of an *adjustable-ratio dc-to-dc transformer*, which is the desired objective of any such converter. The dc voltage transformation ratio M is given by $M = D/D'$, where D is the duty ratio (fractional on-time) of

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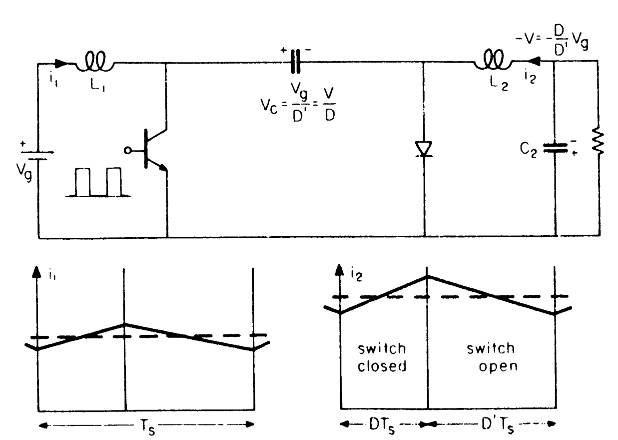

Fig. 1 The original new optimum-topology converter, and the nonpulsating input and output current waveforms.

the transistor switch operated at switching frequency $1/T_s$, and $D' = 1 - D$ is the complementary duty ratio (fractional off-time), when the converter is operated in the continuous inductor-current mode (neither inductor current falls to zero at any time). For a dc input-voltage V_g, the output dc voltage (polarity inverted) is $V = MV_g$. The converter has the same transformation ratio as the conventional buck-boost converter, giving a buck or step-down ratio for $D < 0.5$ and a boost or step-up ratio for $D > 0.5$. The other principal feature is that both the input and output currents are *nonpulsating* (in the continuous inductor current mode), and consist of a dc component with a comparatively small superimposed switching ripple, as also shown in Fig. 1.

Because the input and output inductor currents are essentially constant the switched current is confined entirely within the converter, in the loop formed by the capacitance C, transistor, and diode. When the transistor is off, during the interval $D'T_s$, the input current charges C and the diode carries the sum of the input and output currents; when the transistor is on, during the interval DT_s, the diode is open, the transistor carries the sum of the input and output currents, and C discharges into the load. It may therefore be said that C is a *coupling* or *energy transfer* capacitance, since it stores energy from the input during $D'T_s$ and delivers it to the load during DT_s; this is accomplished by effectively switching C between the input and output circuits. It is easily shown that the average voltage on the coupling capacitance is $V_c = V_g/D' = V/D$.

Furthermore, in the most straightforward design, C is large enough that its voltage ripple is fractionally small so that its voltage is essentially constant at the average value V_c; this result is analogous to that of making the inductances large enough that their respective current ripples are fractionally small.

From yet another point of view, it may be said that energy storage and delivery proceeds *simultaneously* in *two* loops in *both* switch intervals. During DT_s, the input and output loops are closed through the transistor; energy is stored in L_1 from the input, and energy is released from C to L_2 and the load. During $D'T_s$, the input and output loops are closed through the diode: energy is released from the input and L_1 and is stored in C, and simultaneously energy is released from L_2 to support the load. This *symmetry* of the basic new converter is the source of its efficiency advantages and also makes possible several useful extensions [3,4,5] besides those to be introduced here.

3. Development of the isolated version of the new converter

The original new converter of Fig. 1 provides a single, polarity inverted, nonisolated output. For many applications it is essential to provide dc isolation between input and output, and/or multiple outputs of different voltages and polarities.

There is therefore a strong incentive to find a way to introduce an isolation transformer into the original new converter, and the obvious place to do this is somewhere in the inner loop containing the coupling capacitance, transistor, and diode in which the aforementioned switched energy transfer current exists. There are three steps to a simple, elegant solution to this problem.

The first step is to separate the coupling capacitance C into two series capacitances C_a and C_b. The second is to recognize that the connection point between these two capacitances has an indeterminate average or dc voltage, but that this dc voltage can be fixed at zero by connection of an inductance between this point and ground. If the extra inductance is large enough, it diverts a negligible current from that passing through C_a and C_b in series, and so the converter detailed operation is so far unaffected. The third step is merely the separation of the extra inductance into two equal transformer windings, which thus provide the desired dc isolation between input and output.

The result of these three steps is shown in the basic isolated version of the new converter in Fig. 2. With a 1:1 transformer, the voltages and currents in the input and output circuits are the same as in the original nonisolated version. The only difference is that the switched current loop now becomes two loops with equal currents circulating in the same direction.

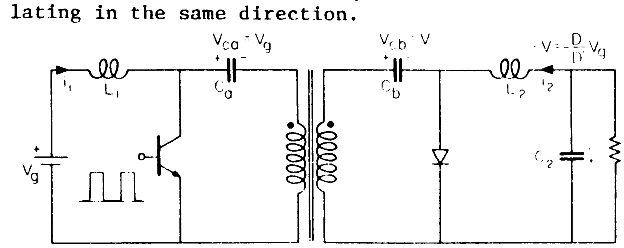

Fig. 2 The new converter with a 1:1 isolation transformer: all the advantages of the original converter are retained.

The salient feature of the isolation method shown in Fig. 2 is that both windings of the transformer are dc blocked by C_a and C_b, and therefore there can be no dc in either winding and so automatic volt-second balance is achieved. Thus, there is no problem of core operating point creep as can occur in push-pull "balanced" isolation arrangements. It follows that, since there can be no average or dc voltage across either transformer winding or either inductance in the circuit of Fig. 2, the voltage on C_a is $V_{ca} = V_g$ and that on C_b is $V_{cb} = V$. It may be noted that $V_{ca} + V_{cb} = V_g + V = V_g/D' = V/D$, the same as the voltage V_c across the single coupling capacitance C in the original converter of Fig. 1.

It is instructive to consider the current paths, voltage distributions, and energy dispositions during the two switch intervals. In Fig. 3(a), conditions are shown during interval $D'T_s$ when the transistor is off. The input current charges L_1 and C_a, and an equal reflected current in the transformer secondary charges C_b. The output inductance L_2 discharges into the load, and the diode carries the sum of the input and output currents. The width of the current path in Fig. 3(a) suggests that the input current i_1 is smaller than the output current i_2, which would be the case for D less than 0.5. Arrows pointing upwards (downwards) indicate elements in which energy is being stored (released). In Fig. 3(b) for interval DT_s when the transistor is on, the input current charges L_1, the reflected discharge current of C_a also discharges C_b and charges L_2 and supplies the load; the transistor again carries the sum of the input and output currents.

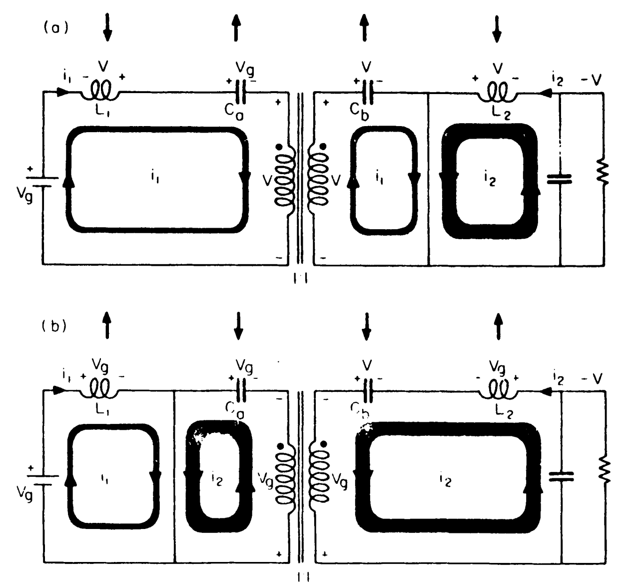

Fig. 3 Current and voltage distributions in the isolated new converter: (a) interval $D'T_s$ when the transistor switch is open; (b) interval DT_s when the switch is closed. Up-pointing arrows indicate energy storage in the adjacent element; down-pointing arrows indicate energy release.

Isolation has thus been achieved in the simplest possible manner by addition only of the necessary transformer (which is single-ended), and the only other modification is separation of the original coupling capacitance into two. Consequently, the configuration of Fig. 2 may be said to represent an optimum-topology dc-isolated new converter.

4. Comparison of the new converter with single-transistor isolated forward and flyback converters

The dc isolated version retains all the features of the original new converter, including a single switching transistor, with a "single-ended" isolation transformer. The compelling simplicity of this circuit immediately invites comparison with familiar single-transistor isolated converters, such as the buck "forward" and the buck-boost "flyback."

Three possible disadvantages of the new converter come to mind. One concerns the two coupling capacitances C_a and C_b: these capacitances transfer the entire power (in ac form) from input to output, and therefore are called upon to handle a substantial ac current. It may appear, therefore, that the esr of these capacitors (or, more directly, their ripple current stress rating) would impose a more severe limitation upon the power handling capacity of the new converter than on, say, the forward converter in which the principal energy transfer is through magnetic rather than electric field energy storage.

The other two possible disadvantages concern the stress levels in the switching transistor and diode. To first order, the stress levels may be defined as the "on" current I_{on}, and the "off" voltage V_{off}. In the new converter, both the transistor and diode carry the sum of the input and output currents, and so perhaps I_{on} is higher than in either the forward or the flyback converters in which the transistor carries only the input current and the diode carries only the output current. In the new converter, which may be viewed as a coalesced boost-buck converter, the series input-inductance L_1 obviously causes the transistor off-voltage to be substantially higher than the input voltage, and so perhaps V_{off} is higher than in either of the other two converters.

We shall see in this section that all three of these conclusions are false: the capacitor ripple current requirements and the transistor and diode stress levels are in fact *the same* in the new converter as in the conventional single-transistor forward and flyback converters. Moreover, there remain other *net advantages* in the new converter, particularly with respect to the size (and losses) of the isolation transformer.

4.1 Capacitor, transistor, and diode comparative stress levels

To compare more quantitatively these converters, let us set up the three circuits to provide the same basic conversion performance. Suppose a one-to-one isolated voltage conversion is required so that $V = V_g$, and consequently (with neglect of losses) the input and output dc currents are each equal to some value I determined by the load resistance. Further, let the switch be driven at a duty ratio $D = 0.5$ in each case.

The three circuits are shown in Figs. 4,5, and 6, together with the transformer primary and secondary voltage and current waveforms v_p, i_p and v_s, i_s appropriate to the chosen operating condition $D = 0.5$. For simplicity it is assumed that the inductances and capacitances are large enough that both current and voltage ripples are negligible, and transistor and diode forward drops are ignored.

In the new converter of Fig. 4, the transformer turns ratio is 1:1 in order to obtain $V = V_g$ with $D = 0.5$; the two coupling capacitances C_a and C_b thus each has the voltage V_g. The relative winding polarity of the transformer secondary is reversed

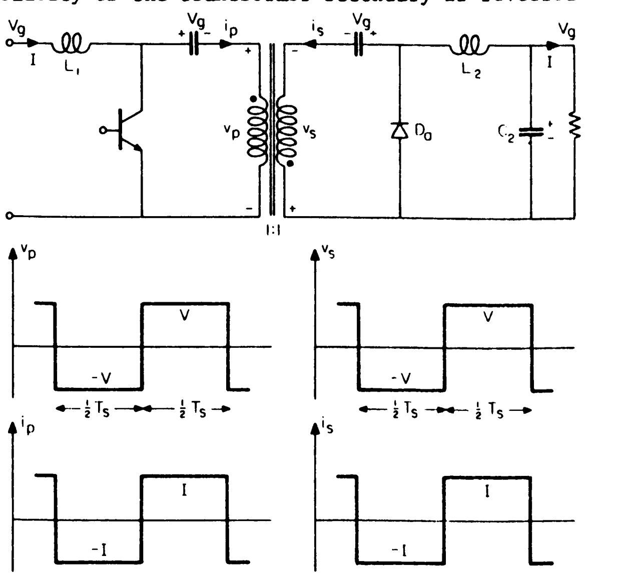

Fig. 4 The new converter with 1:1 isolation transformer, and primary and secondary voltage and current waveforms for $D = 0.5$ for which $V = V_g$.

compared to that of Fig. 2 so that a positive output voltage is obtained as in the corresponding forward converter of Fig. 5, in which the transformer ratio must be 1:2 in order to have $V = V_g$ with $D = 0.5$. The zener diode, necessary for transformer core reset, must have a breakdown voltage V_z of at least $2V_g$ in order that transformer core reset be achieved before the end of the switching cycle; actually, an additional margin would have to be allowed, as shown in the dashed voltage waveforms in Fig. 5. An input filter L_1C_a is included for proper comparison with the new converter of Fig. 4;

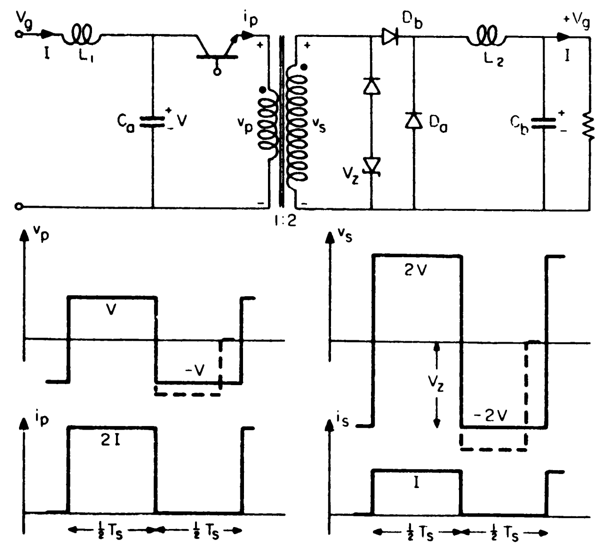

Fig. 5 Transformer voltage and current waveforms in a "forward" converter configured to give $V = V_g$ with $D = 0.5$, for comparison with those of Fig. 4.

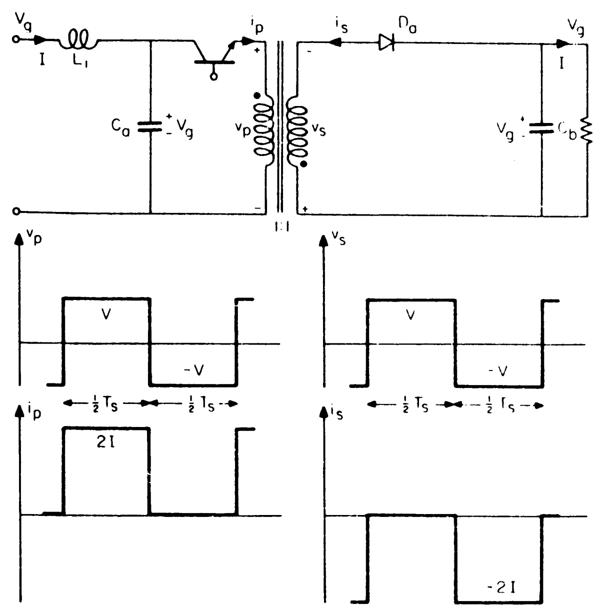

Fig. 6 Transformer voltage and current waveforms in a "flyback" converter configured to give $V = V_g$ with $D = 0.5$, for comparison with those of Fig. 4.

even if the input inductor L_1 were omitted in the forward converter of Fig. 5, the capacitor C_a would still be essential to keep the pulsating input current from being drawn from the V_g supply. In the comparable flyback converter of Fig. 6, the required transformer ratio is again 1:1 so that $V = V_g$ when $D = 0.5$. An input filter L_1C_a is also included to make the input current nonpulsating.

Comparison of the three converters is now easily made by inspection of the circuits and waveforms in Figs. 4, 5, and 6. In the new converter of Fig. 4, each coupling capacitor C_a and C_b carries a square wave current I, and so the ripple current rating requirement is I rms, which is indeed substantial. However, it is seen in Fig. 5 that the current in the input capacitance C_a is $I - i_p$, which also has an rms value of I. The same is true for Fig. 6: each capacitor C_a and C_b carries a ripple current of I rms. Furthermore, in all three circuits, the operating voltage of each capacitor is V_g. Therefore, the *same ripple current rating is required on the capacitors in all three converters*; the only difference is that the forward converter requires one such capacitor whereas the flyback and the new converter each requires two (the output capacitor C_2 in the forward and in the new converter does not have severe ripple current requirements).

It is also easily seen from Figs. 4, 5, and 6 that the transistor in each converter has to pass an on-current of 2 I and has to withstand an off-voltage of 2 V_g. In the forward converter, the off-voltage exceeds 2 V_g if the reset zener has a breakdown voltage greater than 2 V_g. Therefore, the *stress ratings represented by I_{on} and V_{off} for the transistor are the same in all three converters.* Consequently, the transistor dissipation is also the same in all three.

The same result is true for the diodes: the off-voltage V_{off} is 2 V_g in all three circuits; the on-current I_{on} is 2 I for the new converter and for the flyback; for the forward converter, two diodes D_a and D_b are required each to carry an on-current $I_{on} = I$. Consequently, the total diode on-losses are the same in all three converters.

Contrary to the initial impression, therefore, the stress levels on the principal components are the *same* in all three single-transistor isolated converters, and the new converter is at no disadvantage. Let us now consider the design of the isolation transformer itself in each converter.

4.2 Comparative isolation transformer properties

In the new converter of Fig. 4 the isolation transformer has no dc current component in either winding, and leakage inductance can be minimized by use of an ungapped toroid of square-loop material. If the same core and primary winding of resistance R is used in the forward converter of Fig. 5, the secondary will have twice the number of turns of half the wire area to keep the same copper cross-section as in the new converter.

The core and copper are thus set up to be the same in the forward as in the new converter. However, although the core loss is therefore the same, the copper loss in the forward converter is *double* that in the new converter. This occurs because in the primary, the mean square current is twice as large in the forward as in the new converter, and so the $\overline{i_p^2}$ R losses are doubled in the same winding resistance. In the secondary, the mean square current is half as large in the forward converter, but the winding resistance is 4 R because of double the number of turns at half the wire size, so the secondary $\overline{i_s^2}$ R losses are also doubled.

There is a further difference: for use in the forward converter the square-loop core must be *gapped*, since magnetizing current is available in only one direction, and the remanent flux must be reduced to a small value. The effect of this gap is illustrated in Fig. 7. In the forward converter, therefore, the core size must be chosen so that the total flux excursion is not greater than the saturation flux B_s. In contrast, in the new converter, magnetizing current is available in both directions, and so a core that is fully utilized in the forward converter is only half utilized in the new converter. Therefore, a core of half the cross-section could be used in the new converter so that the total flux excursion would be $2B_s$, and as a result the core loss would be halved. The halved area in turn leads to even lower copper loss because the winding lengths are reduced.

Overall, therefore, a smaller, ungapped square-loop core can be used in the new converter than the gapped core necessary in the forward converter, which results in an isolation transformer in the new converter that has lower core loss and lower copper loss. From a general point of view these benefits all stem from the fact that, in the new converter, power is transmitted through the transformer from the input to the output during *both* intervals of the switching cycle, whereas the same average power has to be transmitted during only one interval in the forward converter.

Comparison of the transformer properties in the flyback and new converters shows that the disparity is even more extreme, because in the flyback the core gap must be larger than in the forward converter, as also illustrated in Fig. 7. This is because the transformer is really an inductor, since the transmitted energy is stored in the magnetic field (principally in the air gap) during one interval of the switching cycle and is released to the output

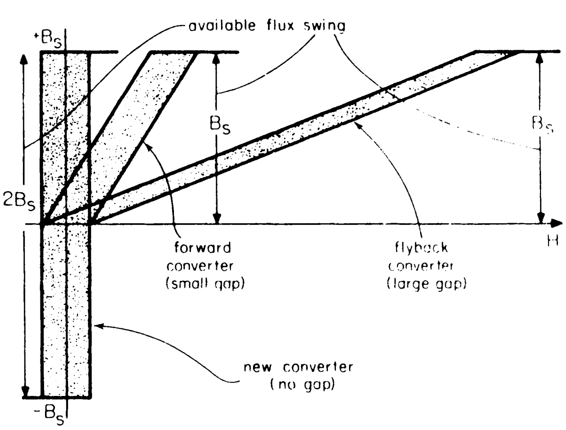

Fig. 7 Comparison of isolation transformer core utilization in the new converter and the forward and flyback converters. Twice the flux swing is available in the new converter.

during the other interval. Consequently, the magnetizing current, which is again available in only one direction, constitutes the total primary or secondary current instead of just a small fraction of it.

4.3 Comparison of the three converters at different operating points

The discussion so far of the comparative properties of the three single-transistor isolated converters has been for one operating condition, $D = 0.5$ that gives $V = V_g$ for all three, since this is a convenient symmetrical case. Comparison of the various stress levels and losses is easily accomplished for other operating conditions, and the results for the original operating point and for two others are assembled in Table 1.

	$V = 0.5V_g$			$V = V_g$			$V = 1.5V_g$		
	new	forward	flyback	new	forward	flyback	new	forward	flyback
D	0.33	0.25	0.33	0.5	0.5	0.5	0.6	0.75	0.6
primary $\overline{i_p^2}R$ loss	$0.5I^2R$	I^2R	$0.75I^2R$	I^2R	$2I^2R$	$2I^2R$	$1.5I^2R$	$3I^2R$	$3.75I^2R$
secondary $\overline{i_s^2}R$ loss	$0.5I^2R$	I^2R	$1.5I^2R$	I^2R	$2I^2R$	$2I^2R$	$1.5I^2R$	$3I^2R$	$2.5I^2R$
C_a ripple $\overline{i_{ca}^2}$	$0.5I^2$	$0.75I^2$	$0.5I^2$	I^2	I^2	I^2	$1.5I^2$	$0.75I^2$	$1.5I^2$
C_b ripple $\overline{i_{cb}^2}$	$0.5I^2$	–	$0.5I^2$	I^2	–	I^2	$1.5I^2$	–	$1.5I^2$
Transistor I_{on}	$1.5I$	$2I$	$1.5I$	$2I$	$2I$	$2I$	$2.5I$	$2I$	$2.5I$
Transistor V_{off}	$1.5V_g$	$2V_g$	$1.5V_g$	$2V_g$	$2V_g$	$2V_g$	$2.5V_g$	$4V_g$	$2.5V_g$
Diode D_a I_{on}	$1.5I$	I	$1.5I$	$2I$	I	$2I$	$2.5I$	I	$2.5I$
Diode D_a V_{off}	$1.5V_g$	$2V_g$	$1.5V_g$	$2V_g$	$2V_g$	$2V_g$	$2.5V_g$	$2V_g$	$2.5V_g$
Diode D_b I_{on}	–	I	–	–	I	–	–	I	–
Diode D_b V_{off}	–	$2V_g$	–	–	$2V_g$	–	–	$6V_g$	–

Table 1. Comparison of capacitor ripple current, transistor, and diode stress levels, and of transformer copper losses, in the three converters of Figs. 4, 5, and 6 operated at three different output voltages.

The comparison conditions are as follows. The three circuits are in Figs. 4, 5, and 6, and the isolation transformer core, for simplicity, is again taken to be the same for all three (except ungapped for the new converter, and appropriately gapped for the other two). The primary winding has the same number of turns of the same wire size for all three, and has a resistance R. Again, the secondary winding is the same as the primary for the new converter and for the flyback, with resistance R, but in the forward converter the secondary has twice the number of turns of half the wire area, and so has a resistance 4 R; thus, the total copper area is the same for all three converters.

Although the transformer turns ratio is selected so that $V = V_g$ for $D = 0.5$ for all three converters, other output voltage settings require a different D for the forward converter than for the other two because of the different effective transformation ratio, as noted in Table 1. The three operating points for which results are given in Table 1 are $V = 0.5\,V_g$, $V = V_g$, and $V = 1.5\,V_g$. In each case, the output current is designated I. For each operation point, the table shows the transformer primary and secondary resistance losses $\overline{i_p^2}\,R$; the mean square ripple currents $\overline{i_{ca}^2}$ and $\overline{i_{cb}^2}$ in the capacitors C_a and C_b; the transistor (first-order) stress levels I_{on} and V_{off}; and the corresponding stress levels in the diodes D_a and D_b.

The center group of results in Table 1, for $V = V_g$ summarizes the results already discussed in detail. The stress levels are the same for all three converters (except that the forward converter has two diodes each carrying half the current of the single diode in the other two converters), and the transformer primary and secondary copper losses are each twice as high in the forward and flyback converters as in the new converter.

In the left-hand group of results in Table 1, for $V = 0.5\,V_g$, it is seen that the transformer losses remain higher in the forward and in the flyback converters, and the disparity is increased in the flyback secondary. The C_a capacitor ripple current is now higher in the forward converter than in the other two, and both the current and voltage stress levels in both the transistor and diode are higher (counting the two diodes together). It is assumed that the reset zener voltage is still $2\,V_g$, the same as for the $D = 0.5$ operating condition.

In the right-hand group of results, for $V = 1.5\,V_g$, the transformer losses remain higher in the forward and flyback converters, and the disparity is increased in the flyback primary. Although the C_a ripple current and the transistor and diode on-currents are smaller in the forward converter than in the other two, the voltage stress levels are considerably higher; this results from the requirement that the reset zener must have a higher breakdown voltage, $6\,V_g$, in order to accomplish core reset in the off-time $0.25\,T_s$. If this same higher breakdown zener were employed in the forward converter operated at lower duty ratios, the voltage stresses would be higher than listed in Table 1.

The conclusion is, therefore, that operation at output voltages other than $V = V_g$ in most respects *increases* the disparity between the new converter and the other two, and so the benefits to be obtained from the new converter configuration become even more striking, particularly when the additional superior features of the transformer design are taken into account.

5. Multiple-output and coupled-inductor extension

Once the isolation transformer has been introduced into the new converter as in Fig. 2, several extensions become obvious. There is no reason why the transformer should be limited to a single 1:1 winding, and multiple outputs of different voltages and polarities are easily obtained from multiple secondary windings, or from a tapped secondary winding as shown in Fig. 8. All of the benefits of the basic new converter are retained in the multiple-output versions; in particular, *all* the output currents *and* the input current are nonpulsating.

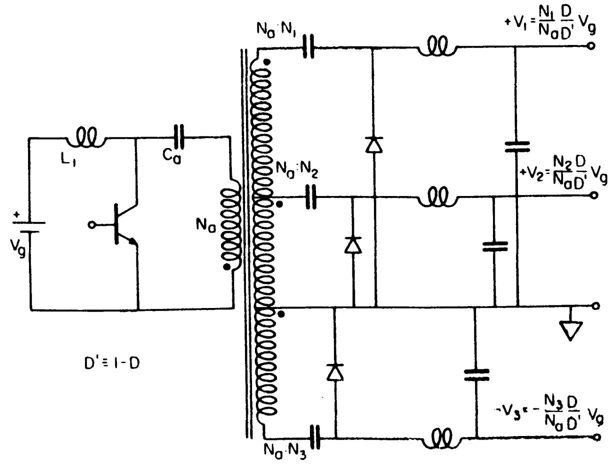

Fig. 8 Extension of the isolated new converter to multiple outputs with arbitrary ratios and polarities.

Another, less obvious, extension involves the possibility of inductor coupling. It has been shown in [3] that the input and output inductors in the basic converter of Fig. 1 can be wound on the same core, with consequent saving in size and weight. Moreover, by judicious selection of the turns ratio and coupling coefficient of the coupled inductors the switching ripple current can be "steered" to either the input or the output circuit, with the result that either the input or output ripple current can be reduced to zero, with obvious performance advantages.

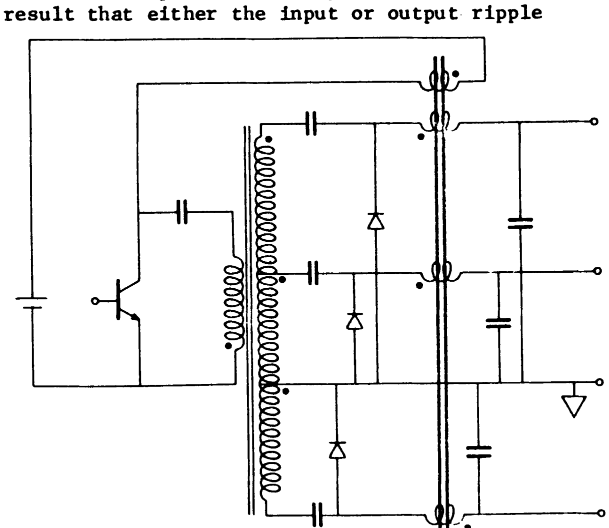

Fig. 9 Any or all of the input and output inductors in the multiple-output new converter can be coupled, which permits the switching current ripple to be "steered" towards or away from a given terminal.

The same opportunity exists in the transformer-isolated multiple-output new converter: any or all of the inductors can be coupled, that is, wound on the same core. Figure 9 shows the same circuit as in Fig. 8 with all of the inductors coupled in this manner, with consequent savings in size and weight. Again, by judicious selection of the turns ratios and coupling coefficients, the ripple currents can be steered to, or away from, the input circuit or any of the outputs.

6. Experimental results, and cross-regulation properties

The test circuit shown in Fig. 10 was constructed with a 1:1:1 isolation transformer, so that the output voltage V_1 is nominally equal to the output voltage V_2. The power switch was operated at 50 kHz with $D \approx 0.5$, and the output voltages were $V_1 \approx V_2 \approx 15$ V. Load currents up to $I_1 = 2$ A and $I_2 = 1$ A were drawn, for a maximum output power of 45 W.

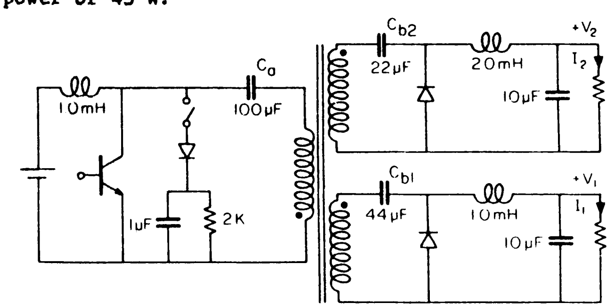

Fig. 10 Test circuit for a two-output isolated new converter, operated at up to 45 W output.

The transformer was designed to take maximum advantage of the low leakage potential. An (ungapped) Magnetics Inc. Square-Permalloy toroid, 51106-2D, was used; the windings were trifilar, each with 39 turns of #26 AWG. The switching frequency of 50 kHz is perhaps rather high for the 2-mil tape thickness, but interest was not centered on core losses in this test circuit. The winding factor is low so that all turns are as close to the core as possible; this results in a leakage inductance of about only 0.3 µH per winding.

The "first-order" off-voltage sustained by the transistor switch is about 37 V. When the circuit was operated at $I_1 = I_2 = 1$ A without the snubber, the additional leakage inductance spike was about 20 V and lasted about 0.12 µsec. With the snubber, the spike was reduced to about 3 V.

One of the important aspects of multiple-output converters is the cross-regulation property. Typically, such a converter is incorporated in a feedback loop in which one output is regulated and the others are "slaved." In this application, the regulated output remains essentially constant, but the slaved output voltages can vary substantially with the currents drawn from all the outputs.

Imperfect cross-regulation in conventional multiple-output converters results from, among other effects, inductor and transformer winding resistance, and unequal diode drops. In the new multiple-output converter, the separate coupling capacitances C_{b1} and C_{b2} in the test circuit of Fig. 10 contribute an additional term to the cross-regulation property because of their unequal discharge during the switch on-interval DT_s. It can easily be shown that the voltage difference $\Delta V_2 - \Delta V_1$ arising from this effect is given by

$$\Delta V_2 - \Delta V_1 = \left(\frac{\Delta I_1}{C_{b1}} - \frac{\Delta I_2}{C_{b2}} \right) \frac{D^2 T_s}{2} \tag{1}$$

Clearly, sufficiently large values of the capacitances C_{b1} and C_{b2} can be used to make the contribution to the cross-regulation from this effect arbitrarily small compared with the remaining effects.

Measurements were made of the cross-regulation and self-regulation properties of the test converter shown in Fig. 10. First, I_1 was varied up to 2 A, while the duty ratio was simultaneously adjusted to keep V_1 constant at 15 V to simulate closed-loop operation with V_1 as the regulated output. Also, I_2 was adjusted to remain at 1 A. The resulting change in V_2 is shown in Fig. 11. The total change ΔV_2 is about 0.9 V for I_1 = 0.2 A to 2 A, or ΔI_1 = 1.8 A. From (1), only about 0.1 V of this change is accounted for by unequal discharge of C_{b1} and C_{b2}; the balance results from series resistance and other parasitic effects.

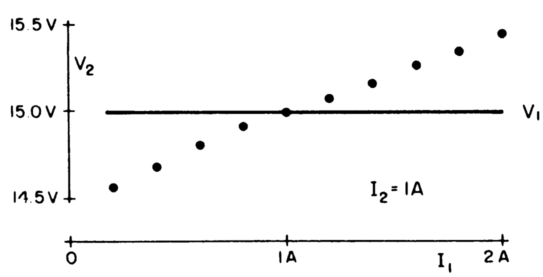

Fig. 11 Cross-regulation property of the circuit of Fig. 10: variation of V_2 as a function of I_1, with V_1 and I_2 maintained constant.

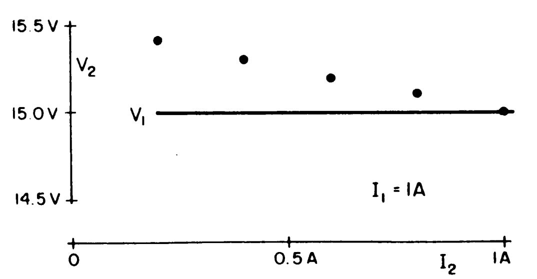

Fig. 12 Self-regulation property of the circuit of Fig. 10: variation of V_2 as a function of I_2, with V_1 and I_1 maintained constant.

Second, I_2 was varied up to 1 A, while the duty ratio was simultaneously adjusted to keep V_1 constant at 15 V, and I_1 was also adjusted to remain at 1 A. The resulting change in V_2 is shown in Fig. 12. The total change ΔV_2 is less than - 0.5 V for ΔI_2 = 0.8 A, of which about a fifth is accounted for by unequal discharge of C_{b1} and C_{b2}.

It is therefore seen that in both the cross-regulation and self-regulation properties the contribution from unequal coupling capacitor charging and discharging is quite small, and is achieved with secondary coupling capacitors of only 44 μF and 22 μF. Larger capacitors, which could easily have been used, would have reduced this effect to negligibility.

7. Conclusion

A recently introduced optimum-topology dc-to-dc switching converter has been extended in a simple and elegant manner to incorporate dc isolation and multiple outputs, with retention of a single switch.

Compared to the conventional single-transistor transformer-isolated forward and flyback converters the new converter has substantial advantages of equal or lower transistor and diode current and voltage stress levels, and also of equal or lower capacitor ripple current stress level. Furthermore, smaller core and winding sizes for the isolation transformer can be employed in the new converter, which also has lower core and copper losses than in the forward and flyback converters. A detailed discussion of these comparisons is given.

The possibility of coupling the input and output inductors, which has previously been shown to lead to reduced, even zero, input or output ripple current in the new converter, is also available in the isolated multiple-output extensions, in which any or all of the input and output inductors can be wound on the same core.

Experimental results are given for a two-output isolated new converter, together with measurements of the cross-regulation properties which are of importance when a multiple-output converter is employed in a feedback loop in which only one output is regulated, as is commonly used in computer power supplies. Work is continuing in all these areas.

Several students in the California Institute of Technology Power Electronics Research Group have participated in various phases of this work. Special acknowledgment is made of the work of graduate student Shi-Ping Hsu, who contributed the test circuits and made the experimental measurements.

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